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A PHASE-AND-GAIN-MATCHED GAIN CONTROL FOR MULTI-CHANNEL DIRECTION FINDING RECEIVERS

N6–ori–71 Task XV
ONR Project No. 076 161
TECHNICAL REPORT NO. 17

ELECTRICAL ENGINEERING RESEARCH LABORATORY
ENGINEERING EXPERIMENT STATION
UNIVERSITY OF ILLINOIS
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Technical Report No. 17

Date
April 1953

Prepared by:

T. R. O'Meara

T. R. O'Meara
Research Associate

Approved by:

H. D. Webb

H. D. Webb
Associate Professor

RADIO DIRECTION FINDING SECTION
ELECTRICAL ENGINEERING RESEARCH LABORATORY
ENGINEERING EXPERIMENT STATION
UNIVERSITY OF ILLINOIS
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ABSTRACT

A method is described for linear operation of pentagrid frequency converter tubes. It is shown how this type of converter operation can be incorporated in a matched gain-control system for a phase-and-gain-matched multi-channel receiver, and how linear operation of the converter tube will reduce the problem of spurious responses in single and double superheterodyne radio receivers.

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I. INTRODUCTION

One of the biggest problems to be faced in multi-channel direction finding receivers is that of finding a satisfactory matched gain control, particularly one suitable for use with an *automatic* gain control system. Either "π" or "T" section step attenuators present a considerable problem in phase and gain matching at frequencies above a few megacycles per second The same is true (perhaps to a lesser extent) of piston type attenuators None of these three types of attenuators is suitable for automatic gain control operation Variations in the dc electrode voltages of vacuum tubes produce gain variations which are very much a function (often badly nonlinear) of the individual tubes, and any matching of gain characteristics must be laboriously repeated with every tube substitution *

Variations in oscillator voltage common to both channels can be used to provide a gain control system which is matched channel to channel provided that the frequency converter stages are operated linearly and certain precautions are observed The basic block diagram for such a gain control scheme is shown in Fig 1 In practice the oscillator output voltage usually *does* vary over the receiver tuning range and hence linear converter operation is much to be recommended for matched dual-channel receivers even if this type of gain control is not contemplated.** In the next section it will be shown that this gain control usually does not require either matching of converter tubes (or other circuit elements) or any point by point adjusting of circuit elements

There is still another problem to be alleviated by linear converter operation in any double superheterodyne receiver, whether intended for direction finder usage or not This problem is that of eliminating spurious responses, independent of any signal at the receiver input, commonly called "birdies" A spurious response often results from the first local oscillator signal, or one of its harmonics, forcing its way through the first intermediate frequency amplifier or power supply cables and beating with a harmonic of the second local oscillator at the second frequency converter to produce a signal output at the second intermediate frequency Since the order of first oscillator signal suppression required to reduce these spurious responses below audibility (or visibility) is often one hundred and twenty decibels or more, it is seen that with usual converter operation the first oscillator shielding and decoupling must be excellent and that the high frequency side responses of the first intermediate frequency tuned circuits (to off frequency signals) must be very small This paper discusses an alternate or supplementary method of attacking the problem of this type of spurious response by linear operation of the frequency converter stages

Still another possible use for gain controls utilizing linear converter operation would be in receivers designed to give low distortion of the modulation envelope of an incoming signal With a gain

*For an example of a matched gain control using variable grid bias see Reference 4.

**In one experimental unit with conventional self bias operation of the converters, a variation in oscillator voltage from eight to fifteen volts was found to produce a (nearly linear) variation of six degrees in the indicated bearing of a direction finder.

control of this type it would not be necessary to use remote cutoff tubes with their variable operating points and their inherent distortion for high signal levels.

II. REDUCTION OF CONVERTER RESPONSE AT HARMONICS OF THE OSCILLATOR SIGNAL

In the normal Class C overdriven operation of frequency converter stages, one can approximate the signal grid transconductance as a function of time (illustrated in Fig. 2b) by a Fourier Series. Let $e_s = E_s \sin(\omega_s t)$ and $e_o = E_o \sin(\omega_o t)$ be the signals applied to the signal and oscillator grids respectively. Following Herold,[1] the plate current (i_p) of a converter tube operated in this fashion can be expressed in terms of this notation as:

$$i_p = g_m(t) E_s \sin \omega_s t \tag{2-1}$$

$$i_p = a_o E_s \sin \omega_s t + E_s \sum_{n=1}^{\infty} a_n \sin \omega_s t \cos n\omega_o t \tag{2-2}$$

where the a_n's are the Fourier coefficients for the expansion of the transconductance wave shape.

$$i_p = a_o E_s \sin \omega_s t + \tfrac{1}{2} E_s \sum_{n=1}^{\infty} a_n \sin (\omega_s + n\omega_o) t$$

$$+ \tfrac{1}{2} E_s \sum_{n=1}^{\infty} a_n \sin (\omega_s - n\omega_o) t \, . \tag{2-3}$$

It is seen that a component at the angular intermediate frequency ω_{IF} is obtained whenever:

$$\pm (\omega_s \pm n\omega_o) = \omega_{IF} \, , \tag{2-4}$$

where n is any integer and the signs are taken to make ω_{IF} positive. Values of ω_s which satisfy Eq. (2-4) will be called the n^{th} harmonic response for a particular n.

For any local oscillator angular frequency ω_1, there is a large number of signal frequencies which will give a response at the receiver intermediate frequency. The ratio of the output at any of the n^{th} harmonic responses to that at the fundamental oscillator frequency is given by a_n (assuming equal input signals at ω_s).

If the oscillator grid voltage is swung over only a linear region of the transconductance characteristic as illustrated in Fig. 2a, a component at the angular intermediate frequency ω_{IF} is obtained only when

$$\pm (\omega_s \pm \omega_o) = \omega_{IF} \qquad (2\text{-}5)$$

In other words, no harmonic responses beyond the first can be obtained. One must pay the price of reduced conversion transconductance in order to obtain linear operation. In practice, converter operation must be somewhat nonlinear, of course, and some weak harmonic responses must result. Table I shows the relative harmonic responses for "linear" and nonlinear operation of the same tube. The magnitudes of the responses vary somewhat from tube to tube, as illustrated by the last two columns. The actual circuits used are illustrated in Fig. 3. The designer should be cautioned that in order to realize the advantages of linear converter operation, the oscillator and buffer circuit outputs must be free from harmonic content. For this reason a tuned circuit was used in the buffer output.

No extensive tests were made with the circuits shown to determine what relative percentage of the harmonic responses resulted from non-linear operation of the frequency changer as compared to harmonic response from harmonic content in the buffer output. However, checks were made by applying the second and third harmonics of the oscillator signal to the buffer grid at the same input level as the fundamental frequency signal. In every case, the converter output was reduced to ten percent or less of the previous output.

Table I indicates that the loss of conversion transconductance entailed by changing from Class C (nonlinear) to linear operation is not great (at least at the oscillator grid signal levels considered here). It is by no means to be suggested that the circuits given here are any sort of optimum with respect to maximum conversion gain for minimum high order harmonic response. The oscillator grid signal levels chosen (one to three volts) were picked on the basis of optimum gain control circuits--to be discussed in a later section.

It is particularly interesting to note (see Table I) that reducing the magnitude of the oscillator signal with non-linear or Class C operation gains very little in reducing the second harmonic response. For example, the gain of the second harmonic response was larger than that of the fundamental response at the one volt oscillator grid level.

TABLE I HARMONIC RESPONSE DATA
RELATIVE DRIVE (COMPARED TO 1750µV), REQUIRED TO GIVE CONSTANT OUTPUT, AT VARIOUS HARMONICS OF THE OSCILLATOR FREQUENCY PLUS OR MINUS THE INTERMEDIATE FREQUENCY (460 kc/s)

Frequencies of Signals Applied to The Signal Grids (Osc Freq.= 3.85mc/s)		Class C Nonlinear Operation (Circuit B, Fig. 3)		Fixed Bias or Linear Converter Operation (Circuit A, Fig. 3)			
		6BA7 Manufacturer No. A		6BA7 Manufacturer No. A		6BA7 Manufacturer No. B	
Harmonic Number	Actual Frequencies mc/s	1 V (RMS) Osc. Grid	3 V (RMS) Osc. Grid	1 V (RMS) Osc. Grid	3 V (RMS) Osc. Grid	1 V (RMS) Osc. Grid	3 V (RMS) Osc. Grid
1st	3.39 4.31	13.1	1	4.2	1.3	4.2	1.4
2nd	7.24 8.16	8.5	14	24.5	3.7	30	4.9
3rd	11.09 12.01	23	5	740	18	800	63
4th	14.94 15.86	46	6	*	340	*	182
5th	18.79 19.71	60	28	*	430	*	560
6th	22.64 23.56	——	17	*	*	*	*

* Indicates No Appreciable Response (Under 5% of Standard Output) For One Volt RMS Signal on the Converter Signal Grid.

If there is a peak in the curve of signal-grid transconductance as a function of oscillator grid voltage, and the operating point is centered near this peak, then the fundamental converter response can become vanishingly small. Of course, such operation is highly undesirable. Figure 4 shows the measured transconductance curve (curve A) for the circuits of Fig. 3 and no peak is indicated. Consequently the second harmonic response data previously mentioned is not so easily explained. If screen and plate voltages are permitted to vary (through a dropping resistor without bypass) then a peaked curve is definitely obtained, and this is illustrated by Curve B of Fig. 4.

III. ANALYSIS OF LINEAR CONVERTER OPERATION

The general sort of plate-current, signal-grid characteristic curve which is required for linear converter operation is shown in Fig. 5a. These characteristic curves are idealized, of course. Unfortunately, no actual tubes with these identical characteristics have been found to exist. However, Fig. 6 shows that some tubes do have characteristics which approximate the ideal characteristic over a reasonable range.

It will be seen that the plate-current, oscillator-grid characteristics, corresponding to the "curves" of Fig. 5a, will also be linear. These data are shown plotted in Fig. 5b. Correspondingly, the curves of signal grid mutual conductance must be linear and this is illustrated for the same idealized tube in Fig. 5c. In fact, assuming linear mutual conductance as a function of oscillator grid voltage and linear plate current characteristics, the general type of curve shown in Fig. 5a is the only type which can result. That is, all of the curves must intersect at a common point, p (Fig. 5a).

Once a linear transconductance curve (Fig. 5c) is assumed, it will be shown why converter tubes which are not matched can be used as "matched" gain control elements. Let

$$e_s = E_s \sin(\omega_s t) \text{ and } e_o = E_o \sin(\omega_o t)$$

be the signals applied to the signal and oscillator grids respectively of two different converter tubes with widely differing (but "linear") transconductance characteristics (see Fig 5c). The linear curve of Fig. 5c gives the following equation for the transconductance of the signal grid (where the constants are for tube 1).

$$g_m = g_{mo}(1 + Ke_o) \tag{3-1}$$

For a second tube the equivalent equation will be

$$g_m' = g_{mo}'(1 + K'e_o) \tag{3-2}$$

The plate current i_p of the first tube will be

$$\begin{aligned} i_p &= e_s\, g_m(t) \\ &= e_s\, g_{mo}(1 + Ke_o) \, . \end{aligned} \tag{3-3}$$

Let the component of the output voltage at the intermediate frequency be e_{IF} (for the first tube).

$$\begin{aligned} e_{IF} &= \tfrac{1}{2}\, g_{mo}\, K\, E_s\, E_o \sin(\omega_s - \omega_o)t \\ &= E_{IF} \sin(\omega_s - \omega_o)t \end{aligned} \tag{3-4}$$

Similarly let the component at the intermediate frequency of the second tube be e'_{IF}

$$\begin{aligned} e_{IF}' &= \tfrac{1}{2}\, g_{mo}'\, K'\, E_s\, E_o \sin(\omega_s - \omega_o)t \\ &= E_{IF}' \sin(\omega_s - \omega_o)t \end{aligned} \tag{3-5}$$

It is seen that the ratio of e_{IF} to e'_{IF} is a constant, independent of either E_s or E_o.*

At this point, the reader may well wonder why the requirement for linear current characteristics exists. The answer hinges upon the desirability of operating both the oscillator and signal grids at fairly large signal levels (on the order of at least one volt RMS) in order to obtain a gain control with a large dynamic range of operation. Assuming nonlinearities which are not matched from one converter to the other, the effective transconductance will vary with signal amplitude and in general will not vary in the same fashion in both channels, thereby creating mismatches.

*In matched channels or receivers (Fig. 1) this ratio can easily be made effectively unity by adjustment of gains in the radio frequency amplifiers. Such correction will hold for all ratios of E_s to E_o.

IV. DYNAMIC RANGE AND OTHER PRACTICAL CONSIDERATIONS OF MATCHED-CHANNEL LINEAR CONVERTER GAIN CONTROLS

This section will give detailed consideration to a linear converter stage applied as a gain control device in a receiver, with the gain variations being produced by varying the oscillator voltage fed to the oscillator grid of the frequency converter tube. More particularly, the remainder of this paper will deal with the application of linear converters to a matched gain control in a dual-channel direction-finding receiver* (Fig. 1). The dynamic range of the receiver will refer to the range of signal grid voltage applied to the signal grid of the converter tube for which the outputs (at the intermediate frequency) of the converter tubes remain constant as a result of varying oscillator grid voltage and with constant ratio of channel A output to channel B output (within prescribed limits).

First let us consider the theoretical dynamic range of operation corresponding to the idealized curves of Fig. 5. The basic limits in such an ideal case will be set by the maximum voltages applicable to the oscillator and signal grids before either cutoff or zero bias is reached In a less ideal case the maximum voltages must confine operation to a reasonably linear region. Let E_{om} and E_{sm} be the maximum peak voltages which can be applied to the oscillator grids and signal grids respectively within these limits Referring to Eq. (3-4), the output voltage E_{IF} is usually set as nearly constant (by optimum driving considerations of the receiver output), depending upon the amount of gain following the converter stage Let us consider a particular "optimum" E_{IF} which shall be called E_{IFO}. Let E_{on} and E_{sn} be the (peak) voltages on the oscillator and signal grids respectively, corresponding to E_{sm} and E_{om}, for a constant peak output voltage E_{IFO}. In other words these are the minimum voltages. The from Eq. (3-5)

$$\left.\begin{aligned} E_{IFO} &= E_{sm}\, E_{on}\, \frac{K}{2}\, g_{mo} \\ E_{IFO} &= E_{sn}\, E_{om}\, \frac{K}{2}\, g_{mo} \end{aligned}\right\} \tag{4-1}$$

The ratio of E_{sm} to E_{sn} (or E_{om} to E_{on}) will give the dynamic gain control range. E_{om} and E_{sm} are known from the linearity requirements on the characteristic curves and E_{on} can be calculated from Eqs. (4-1):

$$E_{on} = \frac{E_{IFO}}{\frac{K}{2}\, g_{mo}\, E_{sm}} \tag{4-2}$$

*The reader who is unfamiliar with these direction finders is referred to Reference 2 and 3.

So that:

$$\frac{E_{om}}{E_{on}} = \frac{K\ g_{mo}\ E_{sm}\ E_{om}}{2E_{IFO}} \qquad (4\text{-}3)$$

Thus the dynamic range is directly proportional to the maximum "undistorted" (or linear) output ability of the converter and inversely proportional to the operating output level of the converter tube. That is, for maximum control range the converter should be near the input of the receiver (as it usually is), so that its operational output and input level is low.

Figure 6 shows the characteristic curves for a 6BA7 type frequency converter tube * The curves approach quite closely to the ideal characteristics represented by the broken lines over a limited operating range In practice, a dynamic range can be obtained which extends into the nonlinear proportions of the characteristics *if certain precautions are observed*.

The worst feature of operating into the nonlinear regions apparently results from shifts in conversion transconductance caused by shifts in operating points. If Class C bias is developed on the oscillator grids this result is obvious. However, cathode resistor bias can also give this effect as changing plate current can cause a shift in operating point and hence in conversion transconductance. The nonlinearities will not in general be the same in both channels and a differential shift in conversion transconductance will result. Cathode bias can be used successfully if the same bias is maintained on both converter tubes by placing a common connection between converter cathodes

Using fixed bias on the converter tubes (as shown in Table I), curves were obtained which are given in Fig. 7 for channel mismatch (bearing error)** as a function of input voltage (to the signal grid of the converter). The circuits which were used are illustrated in Fig. 8 These data were taken with equal input to the two converter tubes, corresponding to a 45° bearing on the bearing indicator oscilloscope Similar data results for differing inputs and differing indicated bearings. For example, the curves for 22½° are similar in character but with slightly less deviation than for 45°.

The various combinations of tubes shown in these curves were picked entirely at random and are quite typical of a wider range of combinations. It is seen that nearly any combination of tubes is satisfactory within a dynamic range of ten but for a maximum dynamic range (10^3) there is something to be gained from selecting tubes--matching is too severe a description of the process. It was found that there was little to be gained *or lost* by adjusting either oscillator or signal grid biases in channel A to match the biases of channel B, but that there *was an optimum range* for both of these bias voltages. At first it was thought on the basis of the characteristic curves (Fig. 6) that one volt signal grid and three volts oscillator grid bias would be about optimum.

*These curves were made on the basis of only a few tubes and hence may be atypical in specific details

**Assuming a 45° bearing, one degree bearing error corresponds to a gain change of about 3.5%, in one channel only.

It was found emperically, however, on the basis of many tests, that two volts signal grid and six volts signal grid bias gave better results *over a 10^3 dynamic range* These values are the ones selected for the data of Fig 7 and the circuits of Fig 8

Phase mismatch was less of a problem with this system than gain mismatch As a rough order of magnitude, about one percent ellipsing* of indicated bearing was obtained for every degree bearing deviation

Figure 9b indicates that equally good results can be obtained for a converter using cathode bias on the signal grid (2 volts) *provided that the same bias is held on both channels* by connecting the cathodes together Figure 9c shows the same data for the same circuits with the exception that the common cathode lead is broken In all cases of Fig 9, additional fixed (battery) bias was maintained on the oscillator grids The circuits corresponding to the data of Fig 9 are illustrated in Fig 10

Some investigations were made with both signal grid and oscillator grid biases obtained by cathode resistors (and common cathode connections) but the results were not as satisfactory as for the cases of external battery bias on the oscillator grid

There are several other practical considerations which warrant considerable attention The signal grid plate current characteristics will not be perfectly linear and if some of the oscillator signal reaches the signal grid by extraneous routes, frequency conversion will occur as a result of this nonlinearity The component of converter output (at the intermediate frequency) resulting from signal grid nonlinearity will "add" to the component resulting from the normal conversion action As the component of oscillator signal at the signal grid must necessarily arrive by extraneous routes (power supply coupling, etc) this component will, in general have a random phase and amplitude in channel A as compared to channel B Thus, the two output signals will have a random shifted magnitude ratio (and a random differential phase shift) as a result of this component It must be remembered that in order to obtain a large dynamic range of gain control at one end of the dynamic range the oscillator signal on the oscillator grid must be very small (considerably less than the signal on the signal grid) Thus although the conversion transconductance resulting from nonlinearities in the signal grid is small indeed, it may still be comparable in magnitude with the conversion transconductance resulting from "normal" converter operation when the oscillator grid voltage is low Signal grid conversion transconductances of magnitudes less than five percent of the oscillator grid conversion transconductances can give considerable channel mismatch (bearing errors) It should be noted that oscillator grid to signal grid coupling** is not a special problem with gain control action . If this coupling is not a problem under normal matched channel operation it will not be for gain control operation, as the signal coupled to the signal grid in this manner will decrease with decreasing oscillator grid signal and decreased gain control settings

*Percent ellipsing is defined as 100 times the ratio of minor to major axis of the ellipse.
**This is usually space charge coupling. See Reference 1

It is worthwhile to caution that the amount of oscillator signal voltage reaching the signal grid of the converter stage will be much greater if the tuned circuits of the radio frequency stages are tuned near the oscillator frequency than if otherwise. Therefore, a double-superheterodyne design for a receiver, with a high first intermediate frequency, is much to be preferred over a single-superheterodyne design with a relatively low intermediate frequency. A double-superheterodyne receiver design also permits the use of two oscillator-converter gain control systems in tandem, giving a larger dynamic range or for the same dynamic range lessening the linearity and shielding requirements.

V. CONCLUSIONS

Linear or near linear operation of a frequency converter stage can be made to greatly reduce the responses of the converter stage to frequencies beating with harmonics of the local oscillator to produce the intermediate frequency. Reducing the harmonic output of the oscillator or reducing the level of the oscillator signal applied to the oscillator grid accomplishes little or nothing towards reducing harmonic responses with conventional Class C converter operation whereas it may accomplish a large reduction for linear operation.

Linear or near linear operation of frequency converter stages can be used with variations in local oscillator amplitude to produce a very satisfactory matched gain control for a phase and gain matched direction-finding receiver. This gain control can be made to operate over a 60 db dynamic range for a single converter tube, with less than one degree bearing error, and requires no matching of components other than a noncritical selection of tubes for optimum results. It is entirely suitable for utilization with an automatic gain control system.

ILLUSTRATIONS

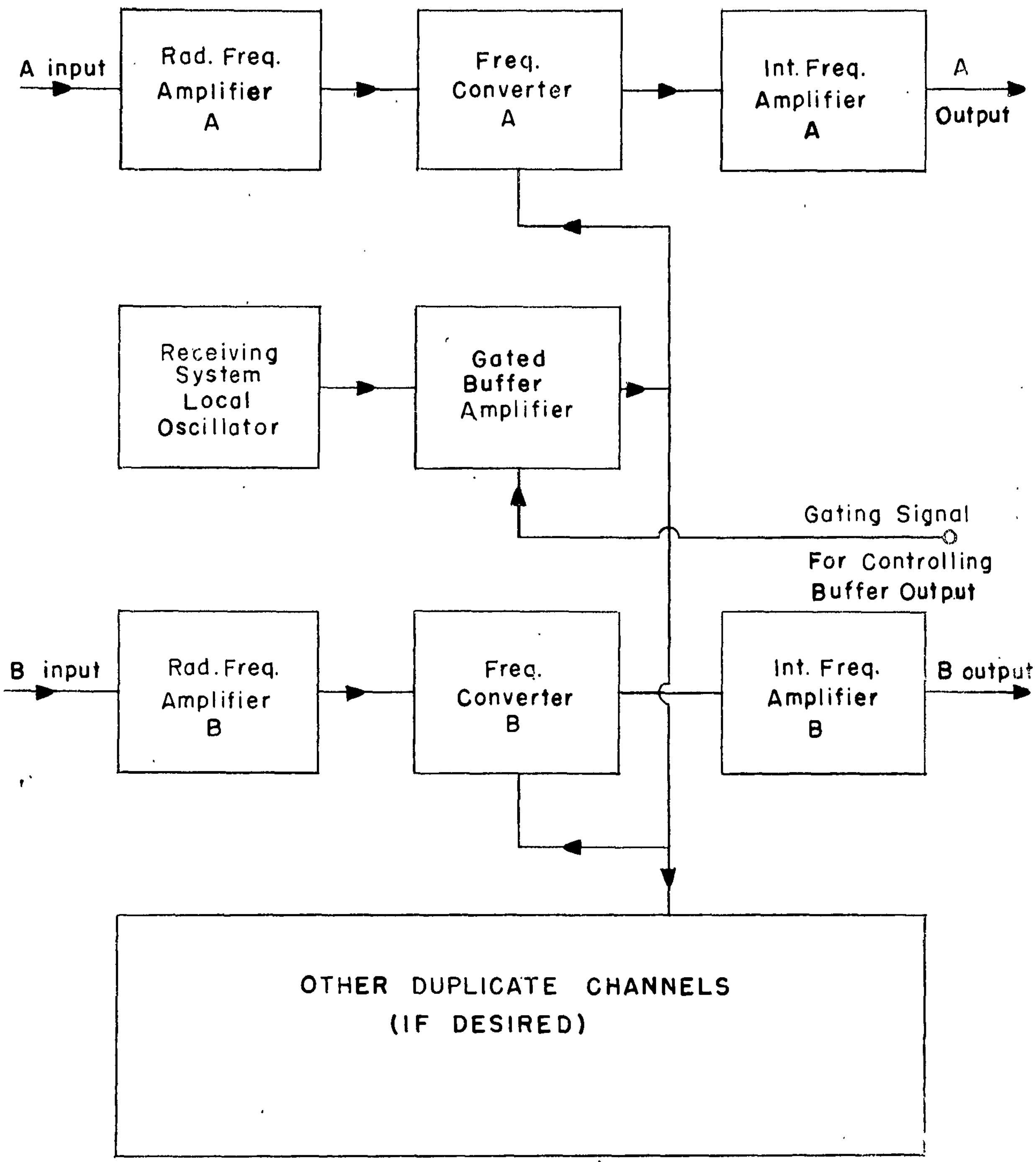

FIGURE 1 BLOCK DIAGRAM OF BASIC GAIN CONTROL SCHEME

Fig. 2a
(Linear Operation)

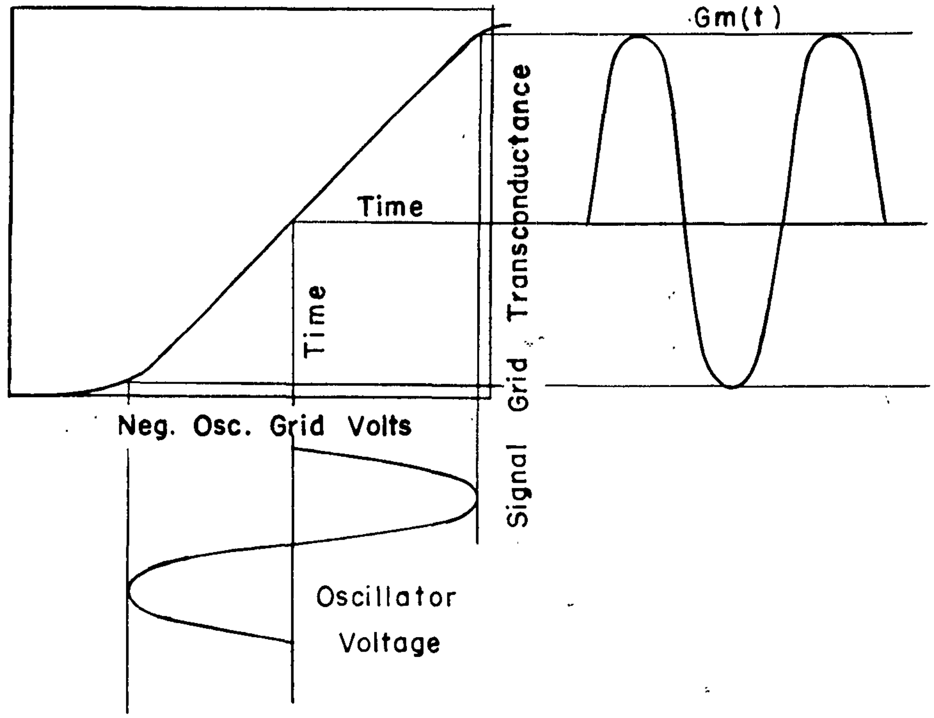

Fig. 2b
(Non-Linear Operation)

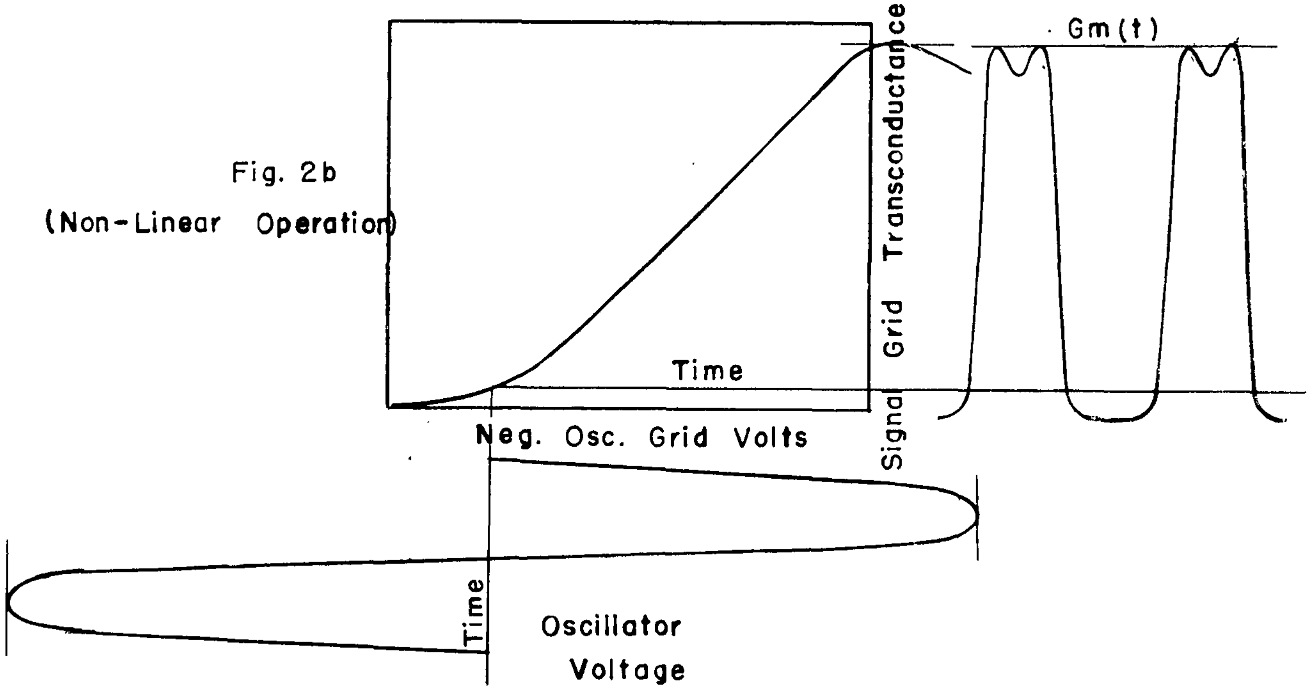

FIGURE 2 TRANSCONDUCTANCE AS A FUNCTION OF OSCILLATOR VOLTAGE FOR LINEAR AND NONLINEAR OPERATION

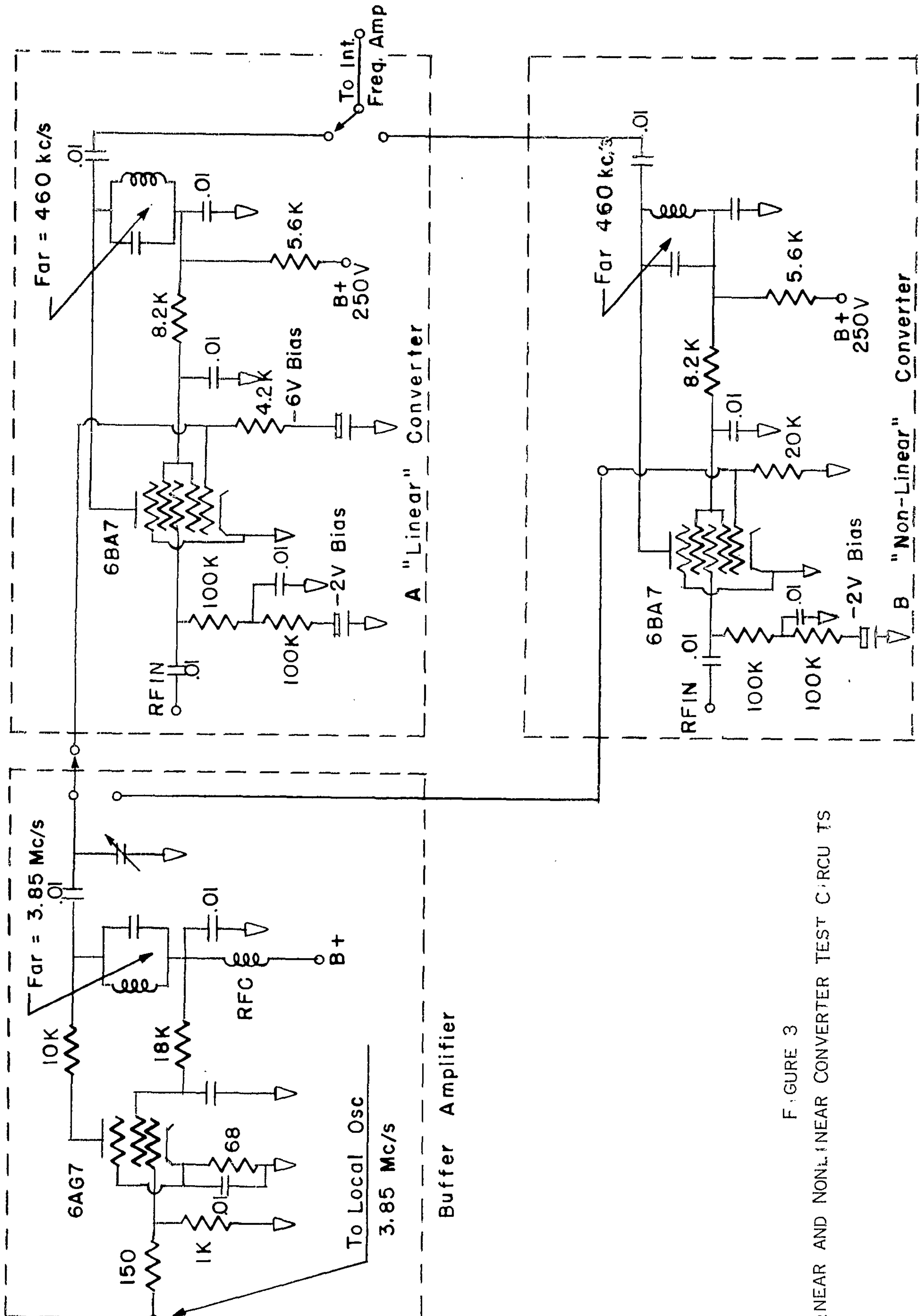

FIGURE 3

LINEAR AND NONLINEAR CONVERTER TEST CIRCUITS

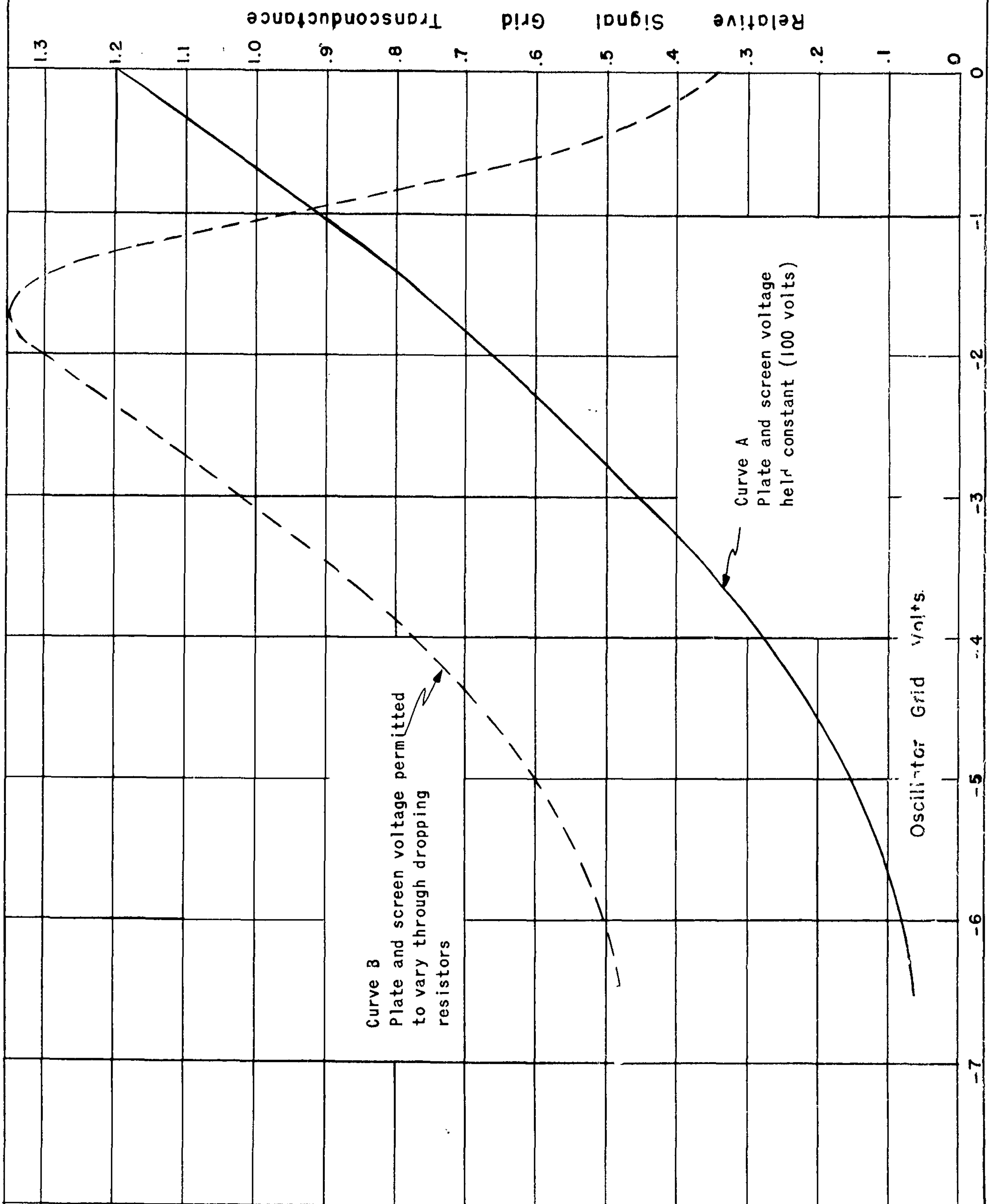

FIGURE 4 SIGNAL GRID TRANSCONDUCTANCE AS A FUNCTION OF OSCILLATOR GRID VOLTAGE (6BA7)

Fig. 5a

Fig. 5b

Fig. 5c

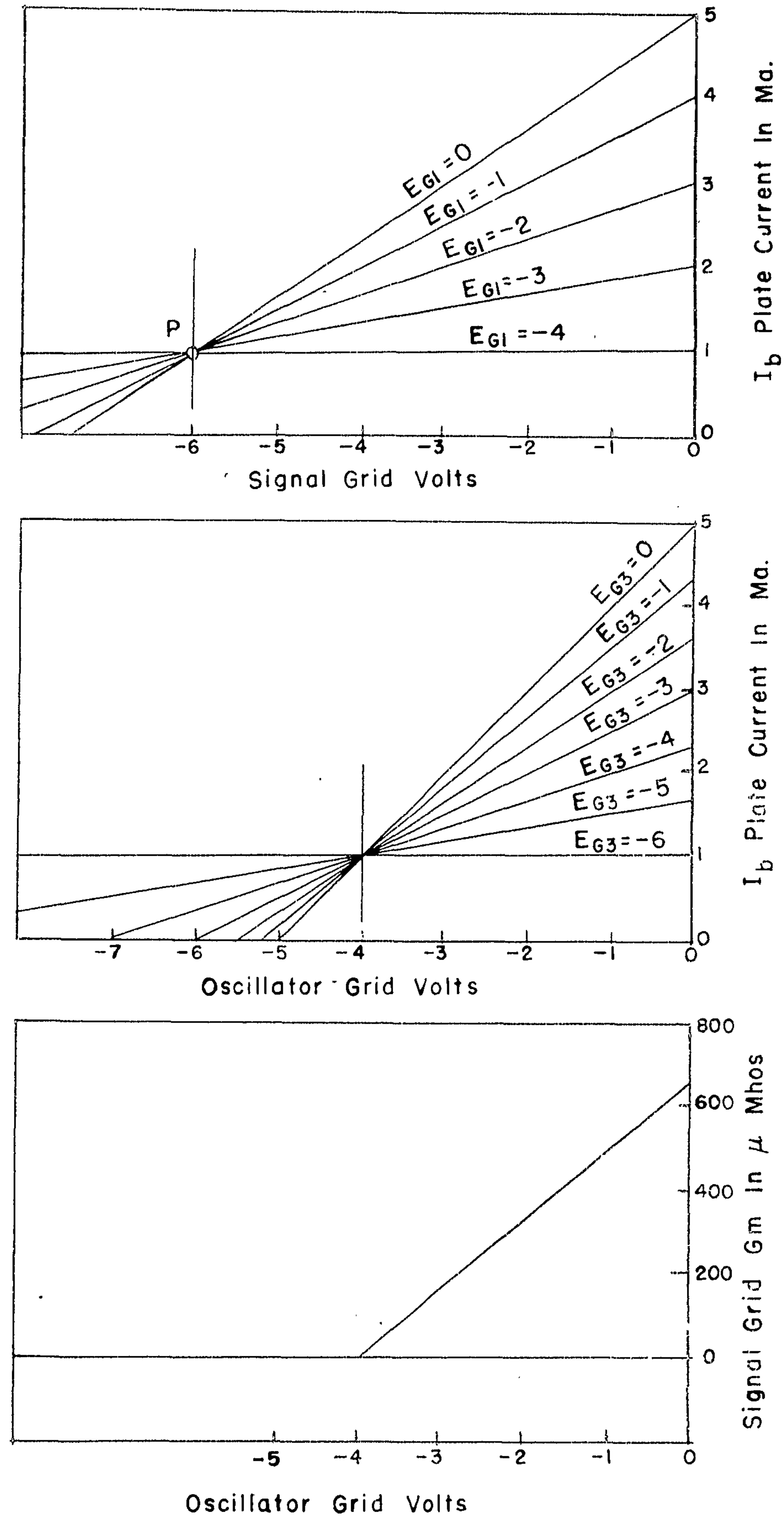

FIGURE 5 IDEALIZED CHARACTERISTIC CURVES FOR LINEAR CONVERTER OPERATION

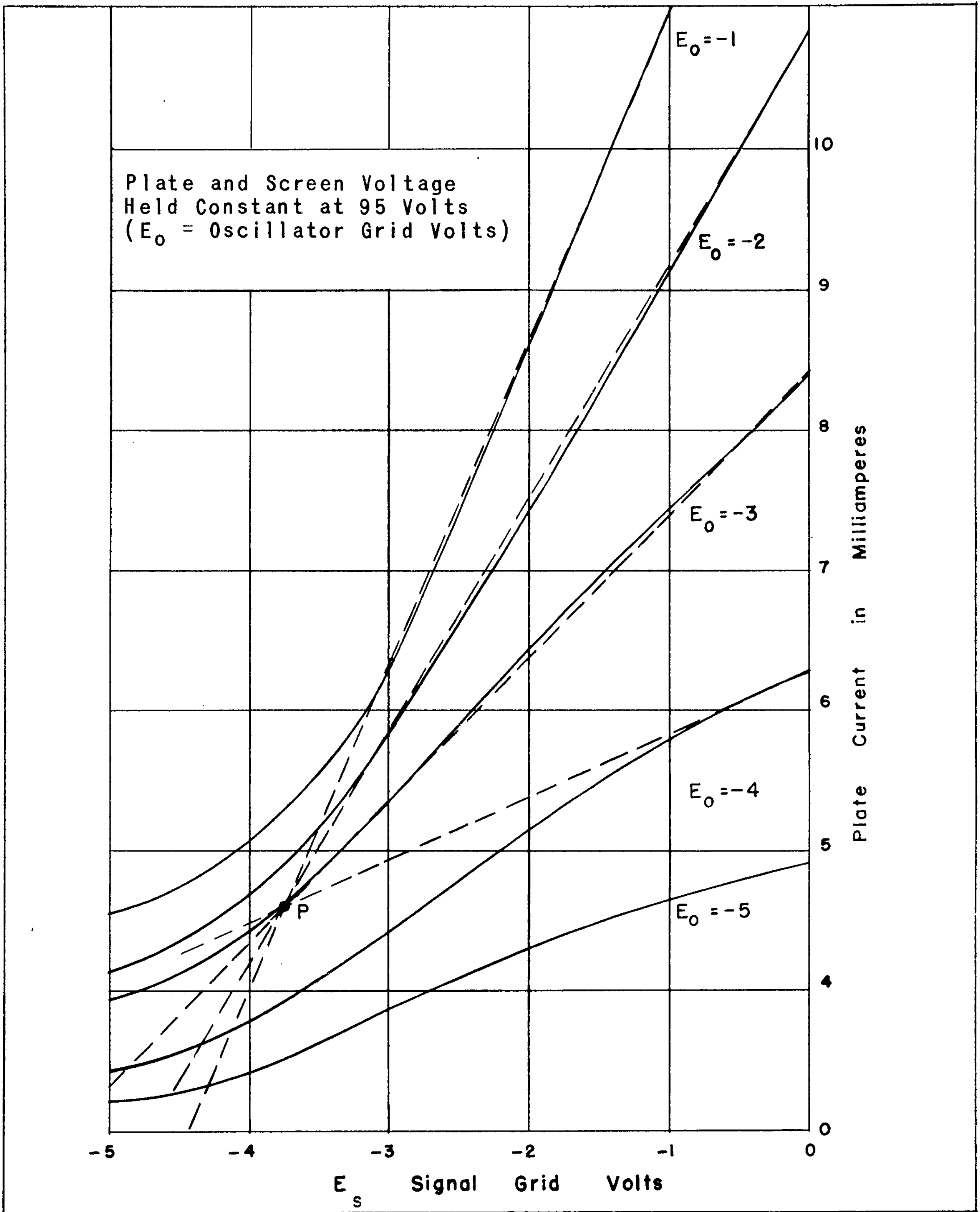

FIGURE 6 PLATE CURRENT AS A FUNCTION OF SIGNAL GRID VOLTS 6BA7

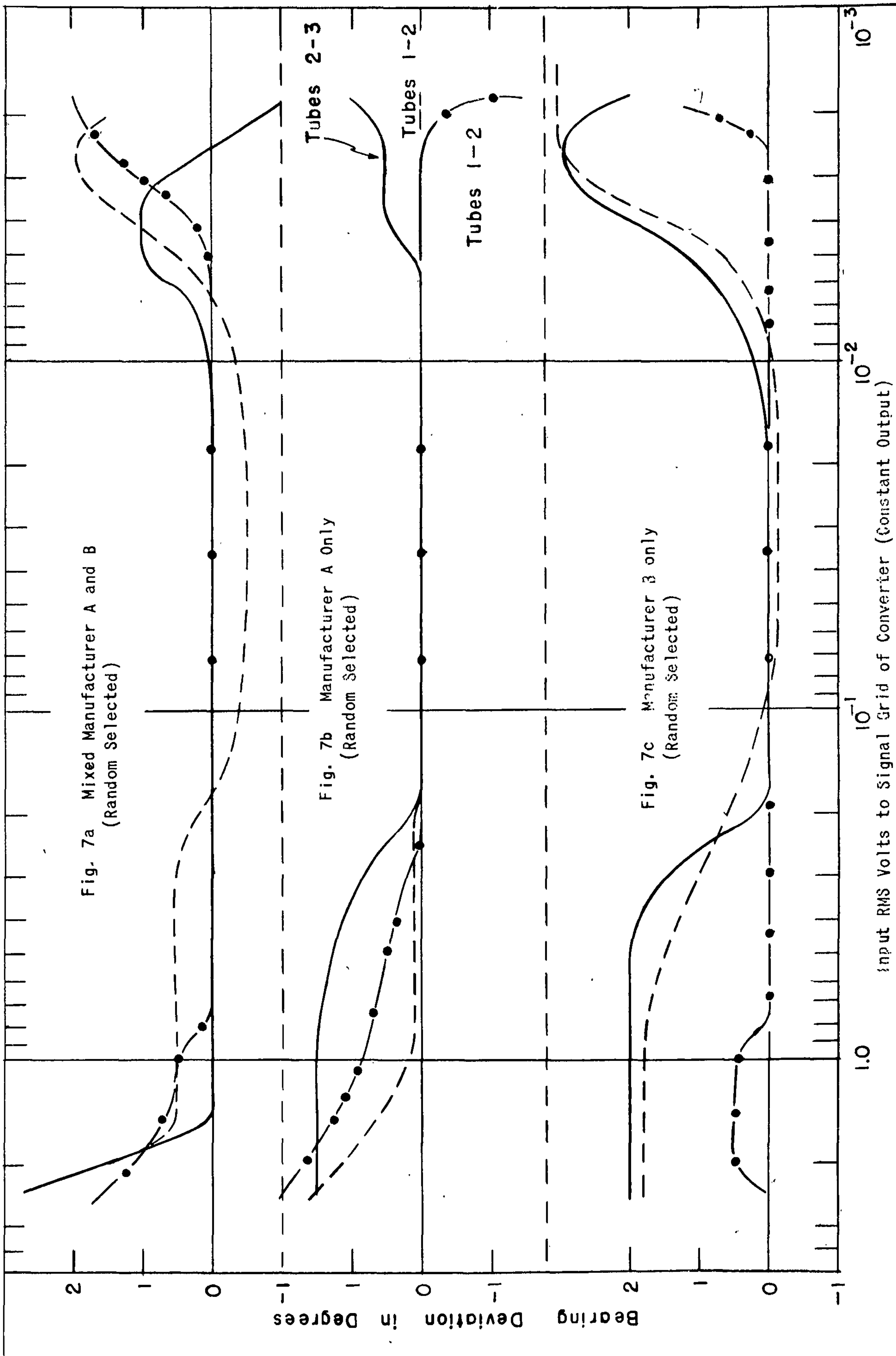

FIGURE 7 GAIN CONTROL OPERATION FOR EXTERNAL BIAS

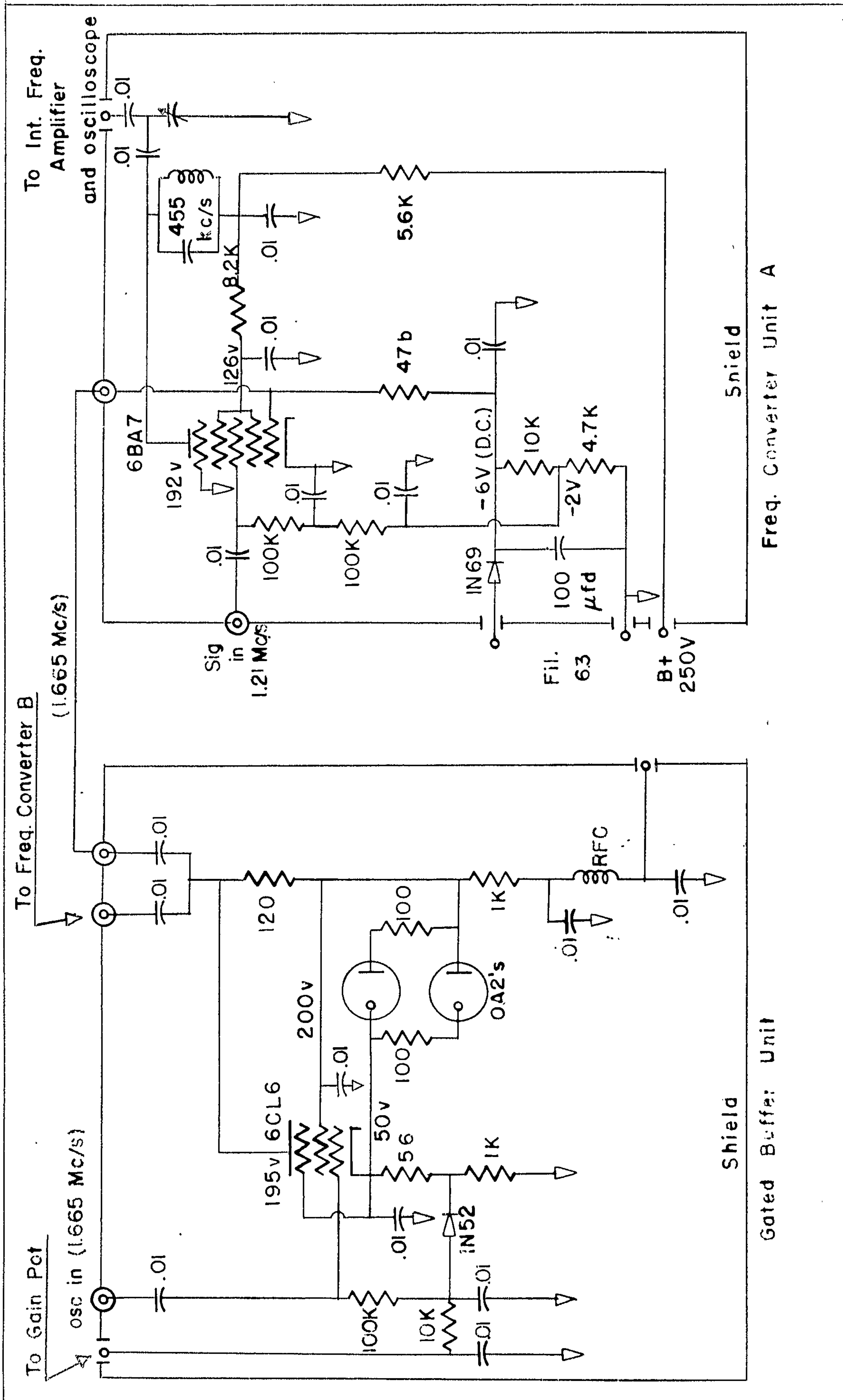

FIGURE 8 GAIN CONTROL SYSTEM WITH FIXED (EXTERNAL) BIASES ON BOTH SIGNAL AND OSCILLATOR GRIDS

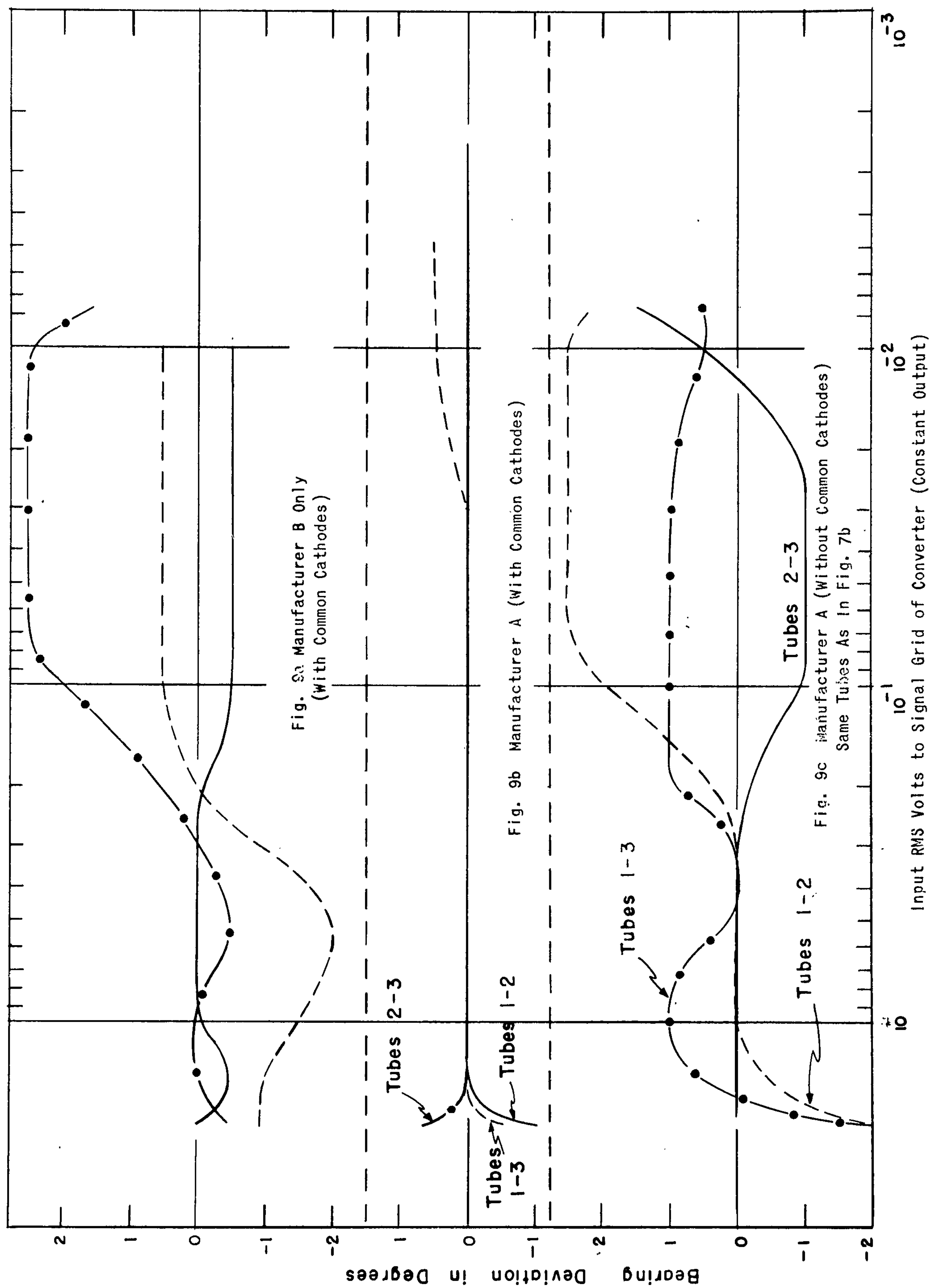

FIGURE 9 GAIN CONTROL OPERATION FOR CATHODE BIAS ON SIGNAL GRID

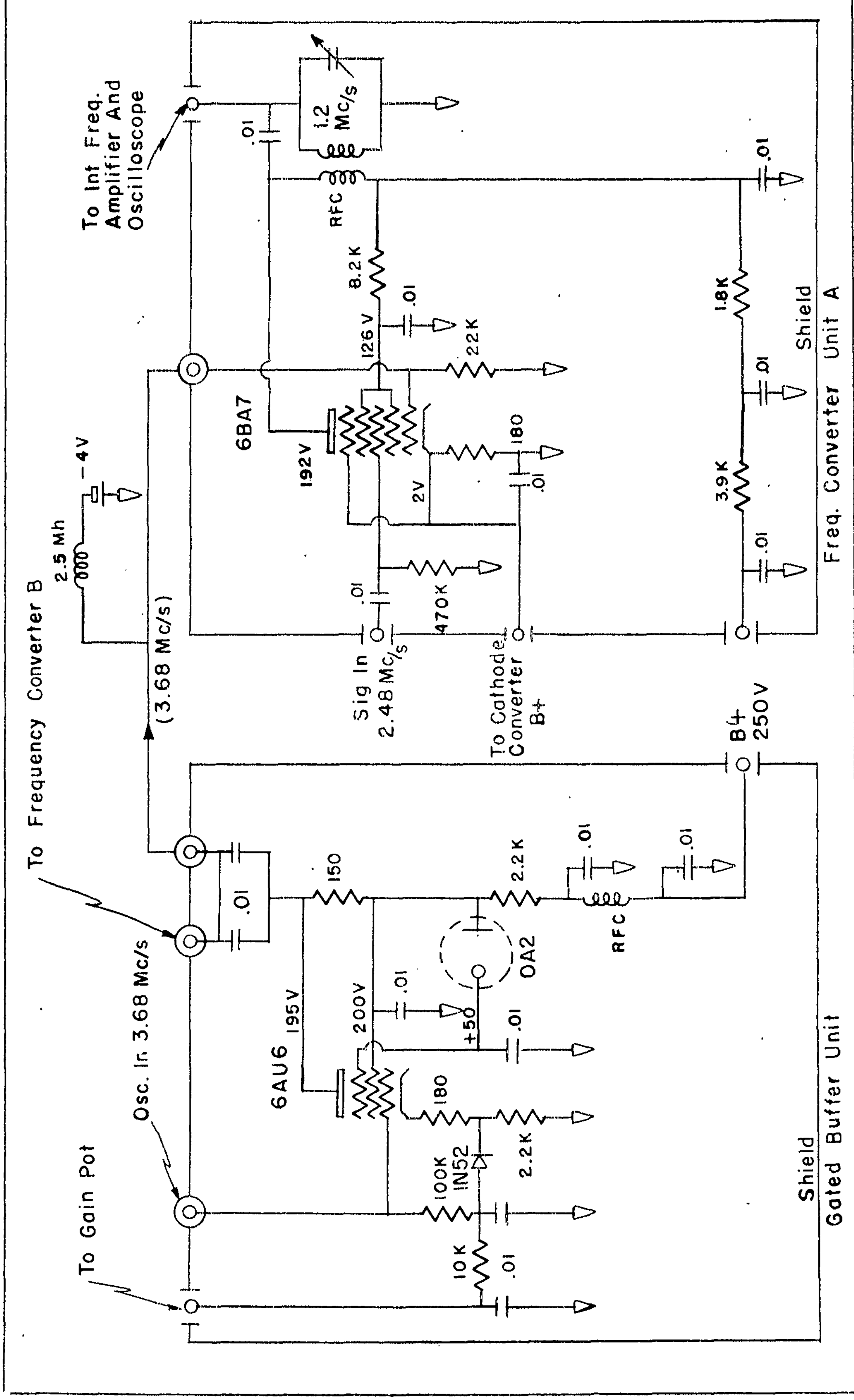

FIGURE 11. GAIN CONTROL SYSTEM WITH CATHODE BIAS ON SIGNAL GRID

Acknowledgement is due L. E. Ernest for his work in testing and constructing the various experimental circuits.

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